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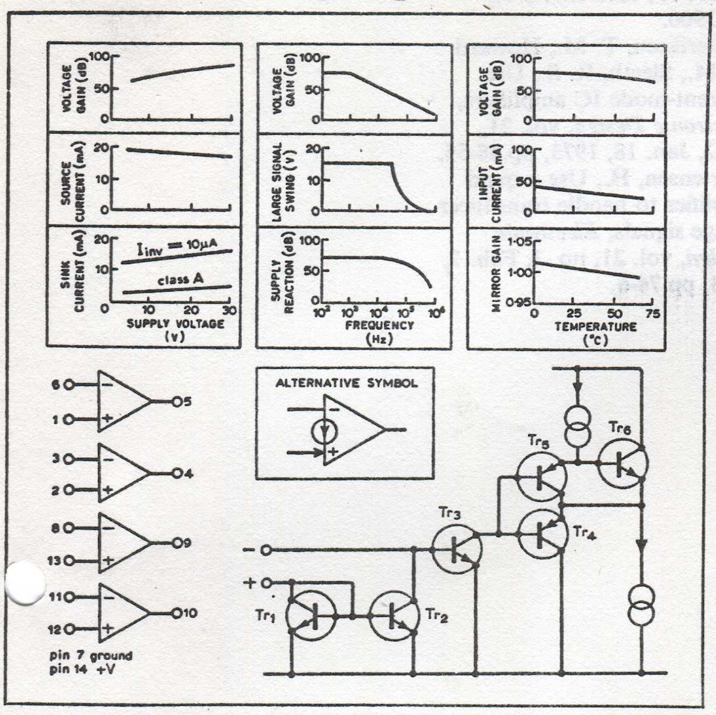
Set 16: c.d.as—signal processing—1

Current differencing amplifiers



Three sets of cards deal with current differencing amplifiers of the LM3900 kind. This set covers signal processing applications, set 17 covers signal generation and a third set deals with various other circuits including test, measurement, detection, logic and driving circuits.

Typical performance
Supply: 15V
R_L: 5kΩ
Voltage gain: 2,800 (69dB)
Output swing: 0.1 to 14.2V
Output current: source 10mA, sink 1.3mA
(overdriving inverting input increases sink current up to >30mA)
Input current: 30nA
Unity-gain bandwidth: 2.5MHz
Slew rate: +0.5V/μs, −20V/μs

N.B. Data is for National Semiconductor LM3900. A similar amplifier is available from Motorola and other manufacturers are expected to "second-source" such devices. Refer to manufacturers data sheets particularly for maximum ratings. While other current-differencing amplifiers may be expected to have similar performance in the circuits to be described it is important that the ratings of particular devices are not exceeded.

Circuit description
Transistors Tr_1, Tr_2 are a current mirror with the collector current of Tr_2, approximately equal to non-

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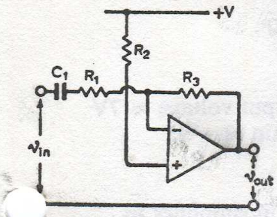
Set 16: c.d.as—signal processing—2

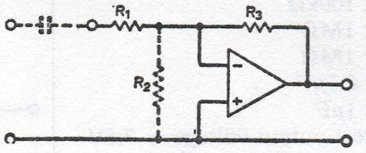
Basic amplifiers—1



Typical performance
Supply: +15V
R_1: 100kΩ
R_2: 2.2MΩ
R_3: 1MΩ
C_1: 0.1μF
Direct output voltage ≈ 7V
Voltage gain ≈ −10

Circuit description—1
No direct current flows in R_1 and hence R_2, R_3 determine the d.c. operating conditions. For perfect balance between the input circuit transistors they will carry equal current and for $V_{be} \ll +V$, the direct output voltage is given by $R_3/R_2 \times (+V)$, i.e. $R_2 = 2R_3$ is the usual condition for maximum available voltage swing with the output biased at supply mid point. As there is no significant alternating current in R_2, it is the alternating currents in R_3 and R_1 that are equal in magnitude while the virtual earth at the inverting input (though ≈0.6V d.c.) gives a voltage gain of $-R_3/R_1$. The addition of reactive components modify the gain, so that a high-frequency roll-off is readily achieved by placing a capacitor across R_3 (corner frequency $1/2\pi R_3C$).

Capacitive coupling may be required to the load, while the reactance of $C_1 \ll R_1$ at lowest frequency. Maximum resistance values of up to 10M may be used, but roll-off due to stray capacitances is likely.



Typical performance
Supply: +15V
R_1: 470kΩ
R_3: 4.7MΩ
R_2: 470kΩ
Direct output voltage ≈ 6V
Voltage gain ≈ −10

Circuit description—2
The base-emitter voltage of the input transistor is ≈0.55V at room temperature, falling by ≈2.5mV for every 1% rise in temperature and by less than this (typically 0.5 to 1mV) for each 1-V increase in the supply voltage. This voltage is thus sufficiently stable to be used as the reference voltage for setting the d.c. output conditions, and the technique may be called the "nV_{be}" biasing method. It is identical in principle to that used in the d.c. feedback pair and the "amplified-diode". If the source has an internal resistance to ground $<R_1$ then direct coupling may be used with R_1 chosen to provide the required input resistance for the circuit and R_3 determining both the voltage gain and the direct output voltage. Resistor R_2 is omitted in this mode as is the input coupling capacitor. Direct output voltage $\approx (R_3/R_1 + 1)V_{be}$. Voltage gain $\approx -R_3/R_1$. The method requires modification both for high and low gains as the direct output voltage may not be convenient. By capacitive coupling to R_1 the d.c. and gain conditions can be made independent, with direct output voltage $\approx (R_3/R_2 + 1)V_{be}$ and voltage gain $\approx -R_3/R_1$.

inverting input current, subtracting from inverting input current at base of Tr_3. The net input current to Tr_3 is (I−) − (I+) and this is amplified by Tr_3 with Tr_5, Tr_6 forming an improved emitter follower output stage. Constant-current generators define the operating conditions while Tr_4 comes into action on over-driving the input to maximize the sink-current. Output depends on the difference between two positive input currents with negative feedback taken to the inverting input when the gain is to be defined. The non-inverting input is outside the feedback loop, and behaves as a forward-biased p-n junction. With resistive negative feedback applied between output and inverting input, the direct currents in the two inputs will be equalized to within the accuracy of the current mirror. If the non-inverting input current is defined by a resistor to +V, the direct output voltage is then a fixed fraction of +V. Transistor Tr_3 base current is ≈30nA, allowing very low bias/signal currents, and like the voltage gain and output current capabilities is controlled over wide temperature and supply variations. An internal regulator (not shown) ensures this by providing the constant currents while also biasing a set of transistors that clamp each input to ≈ −0.3V on negative input swings.

Further reading

Frederiksen, T. M., Howard, W. M., Sleeth, R. S., The LM3900—A New Current-Differencing Quad of ±Input Amplifiers, National Semiconductor application note AN72.

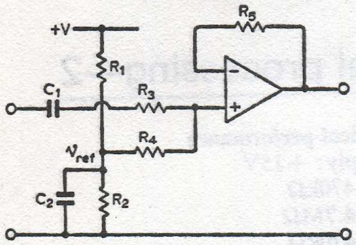
Frederiksen, T. M., Norton quad amplifier subtracts from costs, adds to design options, *Electronics*, Dec. 6, 1973, pp.116-20.

Motorola Linear Integrated Circuit Data Book, pp.7-446, 7-453, 7-456 and 7-463; data sheets on MC3301P and MC3401P amplifiers.

National Semiconductor, Linear Integrated Circuits, pp.226-33, data sheets on LM3900.

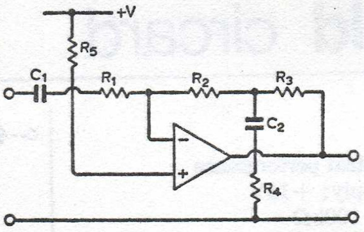
Frederiksen, T. M., Howard, W. M., Sleeth, R. S., Use Current-mode IC amplifiers, *Electronic Design*, vol. 21, no. 2, Jan. 18, 1973, pp.48-55.

Mortensen, H., Use a quad amplifier to handle transducer bridge signals, *Electronic Design*, vol. 21, no. 3, Feb. 1, 1973, pp.74-6.



Typical performance
Supply: +15V
R_1: 47kΩ
R_2: 47kΩ
R_3: 100kΩ
R_4: 1MΩ
R_5: 1MΩ
C_1: 0.1μF
C_2: 1μF
Direct output voltage ≈ 7.5V
Voltage gain ≈ +9.5

Circuit description—3
A third variation on the biasing methods available, is to take the bias resistor R_L to a separate reference voltage V_{ref} which can be decoupled to make the output voltage much less dependent on supply ripple. A single reference voltage (here equal to $+V/2$) may be used for a number of separate amplifiers separate control of the quiescent output conditions is by variation of R_4 for each amplifier while adjustment of R_1, R_2 varies all of them simultaneously. The amplifier is shown with the signal applied to the non-inverting input. No feedback is available at this input and so the impedance of the input transistor affects the input current. At room temperatures, $r_1 \approx 0.026/I_{R4}$, giving a value >3kΩ for the values shown. This reduces the gain to about 3% below the simple theoretical relationship R_5/R_3. This biasing method is equally applicable to the inverting amplifiers.



Typical performance
Supply: +15V
R_1: 1MΩ
R_2: 1MΩ
R_3: 1MΩ
R_4: 10kΩ
R_5: 2.2MΩ
C_1: 0.1μF
C_2: 4.7μF
Direct output voltage ≈ 7V
Voltage gain ≈ −95

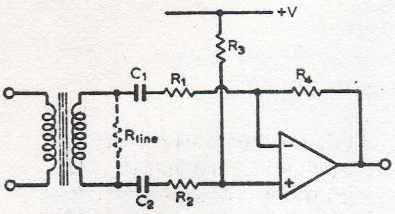
Circuit description—4
The value of the feedback resistor is limited to a few megohms for several reasons (bias instability, effect of stray capacitance, noise and hum). If it is required to have a high input resistance and high voltage gain than the a.c. and d.c. feedback must be different. As shown, R_4 and R_3 constitute a potential divider for the output signal while only R_3 is involved in the d.c. feedback. Output voltage has a quiescent value of $+V(R_2+R_3)/R_5$. Voltage gain is $\approx(-R_2/R_1)(R_3/R_4+1)$. Where $R_2=R_1$, a convenient condition, the voltage gain simplifies to $-(R_3/R_4+1)$. However when the ratio R_3/R_4 is large the feedback theory demands that the limited open-loop gain be taken into account. In practice, a ratio that should set the gain to −20 will do so to within about 1%, while a nominal gain of −100 would be nearer to −95.

Cross references

Set 5, cards 5, 8, 9, 10.
Set 7, cards 4, 10.
Set 10, cards 1, 9.
Set 16, cards 2, 5, 6, 10.

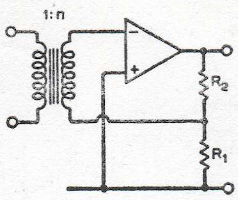
wireless world circard Set 16: c.d.as—signal processing—3

Basic amplifiers—2



Typical performance
Supply: +15V
R_1, R_2: 10kΩ
R_3: 220kΩ
R_4: 100kΩ
C_1: 1μF
C_2: 1μF
Direct output voltage ≈ 7V
Voltage gain ≈ −10
(transformer secondary output)

Circuit description—1
Common-mode signals are a problem when transmission over lines has to take place in a noisy environment. By coupling the signals through a transformer such common mode signals are minimized, while the anti-phase inputs of the current differencing amplifier offer a further improvement. Any common-mode voltage at the transformer secondary produces equal currents at the two inputs largely cancelling each other because the gain at the two inputs is equal and opposite. R_{line} is inserted to achieve the correct loading on the source with R_1, R_L sufficiently larger not to affect that loading. Quiescent output voltage ≈ $V(R_3/R_4)$; voltage gain ≈ $-R_4/R_1$.



Typical performance
Supply: +15V
R_1: 10kΩ
R_2: 100kΩ
Direct output voltage ≈ 6V
Voltage gain ≈ 11
(transformer secondary to output)

Circuit description—2
Where the source is inductive or is to be transformer-coupled, a variant of the "nV_{be}" biasing method provides a simple solution. Again there is the restriction that the direct output voltage and the gain of the amplifier itself are controlled by the same resistor ratio but decoupling part of R_1 to ground can make the ratio for signal frequencies ≫ the ratio at d.c. if required. In the extreme case, R_1 can be completely decoupled giving the full open-loop gain of the amplifier. This coupled with the step-up turns ratio of the transformer gives a very high overall gain. As shown, the overall voltage gain is ≈ $n(R_2/R_1+1)$ and the quiescent output voltage is ≈ $(R_2/R_1+1)V_{be}$. Because the input current required by the amplifier is very small (≪ 1μA) the effective input impedance remains high regardless of the gain, and high step-up ratios are possible. This yields a very sensitive microphone amplifier though the noise performance is unlikely to allow use in audio applications.

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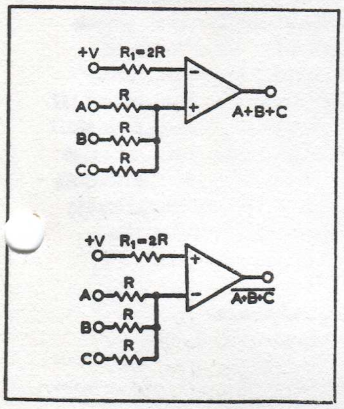
Logic gates



Typical performance
Supply: +20V
IC: ¼ LM3900
R_1: 150kΩ
R: 82kΩ
Output logic 0: 150mV
Output logic 1: 19.2V
For inputs commoned, output changes state for input voltage ≈ 20% of +V.

Circuit description
Availability of two current inputs simplifies the design of basic logic gates with these amplifiers. For example, a low current at one input can hold the output in one desired state while the other input receives the sum of the currents from two or more inputs. This sum can be set to overcome the bias when only one input is high or only if all are simultaneously high, leading to OR and AND-type circuits respectively. (The amplifier is working as a high-gain comparator and can also provide a majority-gate in which any two out of three inputs are enough to provide the required output. By extension some of the simpler forms of threshold logic are possible by scaling the values of resistors to assign a different weight to their importance in decision making.) In the first configuration, if any input is high the current driven into the non-inverting input exceeds the inverting input current and the output is driven high, i.e. an OR gate. The remaining input resistors connected to logic 0 bypass a small portion of that current ($<0.5V/R$ for each resistor) but unless the number of inputs is large and/or the supply voltage is low, this is not a problem. Speed of response is limited to ≈0.5V/μs for positive swings and up to 20V/μs for negative swings though the fall in voltage is slower as logic 0 is approached. By interchanging the inverting and non-inverting inputs with no change in component values, a NOR gate is produced. This flexibility of being able to produce different logic functions from the same package is very attractive. In addition, one or more of the amplifiers can be used to provide astable, Schmitt trigger functions, etc. for obtaining the appropriate waveforms with which to drive the gates.

Component changes
+V Normal voltage range is +4 to +36V, but some devices will operate to <3V without difficulty.
R_1, R Ratio of these resistances is chosen to ensure that with the lowest expected value of logic 1 to any one input that, the resulting current flow into the non-inverting input is sufficient to overcome

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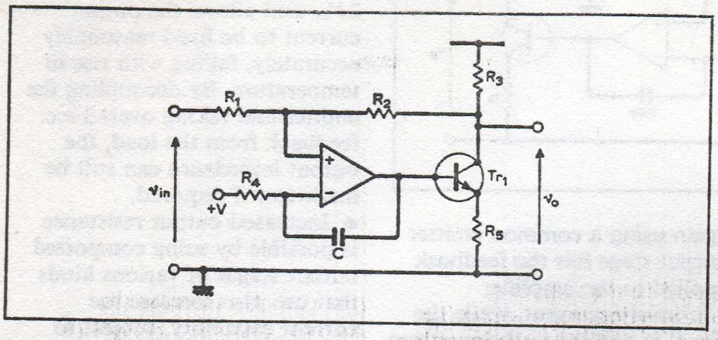
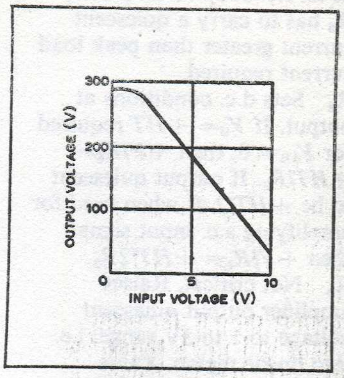
Set 16: c.d.as—signal processing—5

High voltage amplifiers



Typical performance
Supply: +15V, +300V (HT)
R_1: 330kΩ
R_2: 10MΩ
R_3, R_4: 470kΩ
R_5: 1kΩ
C_1: 100pF
Voltage gain: −31.2 (d.c. to 1kHz)
Cut-off frequency: 3kHz
Output impedance: <10kΩ (d.c. to 1kHz)
Input impedance: 330kΩ (d.c. to 1kHz)



Circuit description
Where a high output voltage swing is required the amplifier must be fed from a separate low-voltage supply, and a suitable high-voltage transistor employed to withstand the main supply voltage. The configuration depends on whether it is the output voltage or current that is to be defined. If the former, then the feedback is taken from in shunt with the load. For an inverting-gain amplifier and the transistor in the common-emitter mode, the inverting gain of the transistor necessitates that the feedback be applied to what is normally considered as the non-inverting input. For an input of 0V d.c., the current flow in R_1 is small and that in R_2 is forced by the feedback to equal that in R_4. If in this condition the output is desired to be $+HT$ then $+HT/R_2 = +V/R_4$ gives the required value of R_4. Overall voltage gain is given by $(-R_2/R_1)$ as the circuit is effectively a "see-saw" amplifier while the input resistance is approximately R_1. Resistance R_5 introduces a small amount of negative feedback into the output stage and raises the amplifier quiescent voltage well into its linear region. Capacitor C_1 modifies the gain/frequency characteristic to maintain stability at the higher loop gain. Output voltage swing can be up to 95% of the supply voltage if lightly loaded and the negative feedback keeps the output impedance reasonably low.

Component changes
R_1, R_2 These set the input resistance and the voltage gain $R_2 \gg R_3$ to minimize loading on

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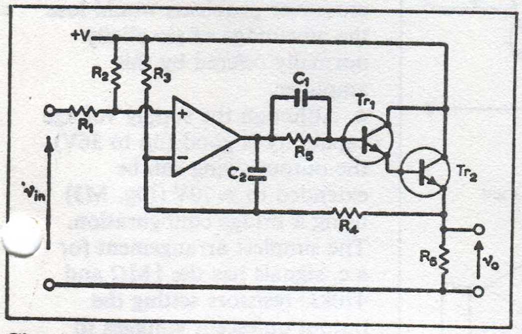
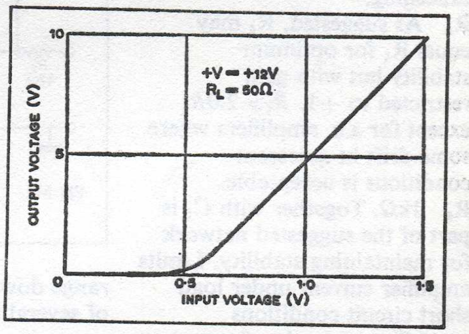
Set 16: c.d.as—signal processing—6

Power amplifiers



Typical performance
Supply: +10V
R_1: 100kΩ
R_2, R_3, R_4: 1MΩ
R_5: 1kΩ
R_6: 50Ω
C_1: 1nF
C_2: 10pF
Tr_1: BFR41
Tr_2: TIP3055
V_1: 1.03V
V_o: 5.0V
Voltage gain: 9.7
Change in V for
I_L of 0-1A 0.05%
V_s of 10-14V 0.5%



Circuit description
Addition of a Darlington-connected pair of transistors increases the output current capability from 10mA to 1·5A depending on the ratings of the transistors used. One restriction is that the V_{be}'s of the transistors limit the output voltage to around 2.5V below the supply level allowing for the amplifier internal saturation. The additional phase shifts that may occur even in an emitter follower make external compensation desirable. For supply, input and output to be all positive, the configuration shown is adequate, where with $R_2 = R_3$, V_o varies linearly with V_1 provided V_1 is above the amplifier internal V_{be}. The relationship is then $V_o \approx V_{be} + (R_4/R_1)(V_1 - V_{be})$ as the currents in R_1 and R_4 have to be equal. This means that as a d.c. amplifier it is of relatively low accuracy but is quite suitable for supplying small d.c. motors under the control of a phase-locked loop. A combination of the techniques for increasing the current ratings and voltage could allow the production of high-power amplifiers. Replacing the emitter follower stages by common emitter amplifier increases the available positive output swing to within a hundred millivolts of the positive supply.

Component changes
R_1 This sets the voltage gain in conjunction with R_4. Because of the V_{be} offset, the output voltage becomes temperature dependent particularly for V_1 comparable to V_{be}, i.e. high gains are not compatible with good stability in this configuration. R_1 100k to 10MΩ.
R_2, R_3 These provide forward bias for each of the inputs allowing the output to be

the output. R_2 1M to 22MΩ.
R_3 Load resistance, dictated by user requirements. For use as an r.c.-coupled amplifier, R_3 has to carry a quiescent current greater than peak load current required.
R_4 Sets d.c. conditions at output. If $V_o = +HT$ required for $V_{in} = 0$, then $+V/R_4 = +HT/R_2$. If output quiescent to be $+HT/2$ as when used for amplifying a.c. input signal, then $+V/R_4 = +HT/2R_2$.
R_5 Not critical. Raises amplifier output quiescent voltage to 1 to 3V range, i.e. into linear region. Value dependent on output quiescent current but might be 50Ω to 5kΩ.
+V Chosen to suit amplifier, and available supplies (+4 to +36V).
+HT Dictated by load requirements.
Tr_1 Must have voltage rating in excess of +HT particularly if inductive loading possible.

Circuit modifications

- A non-inverting amplifier again using a common emitter output stage has the feedback applied to the amplifier non-inverting input, while the signal is applied to the inverting input (the inversion in the transistor ensuring the correct overall sign for the gain). For a transfer function that passes through the origin a pre-biasing network is required. Again the gain is set by the ratio of the feedback to the input resistor.
- For a high output-impedance, the feedback is taken from the transistor emitter. The method can be employed where the load is to be transformer coupled and the direct voltage drop across the primary is too small to allow of the d.c. feedback. The two 10-MΩ resistors define the potential at the transistor emitter as about $2V_{be}$ and allows the output current to be fixed reasonably accurately, falling with rise in temperature. By decoupling the emitter and taking overall a.c. feedback from the load, the output impedance can still be made low if required.
- Increased output resistance is possible by using compound output stages of various kinds that can also increase the current capability, subject to device power limitations. An f.e.t. draws no current from the amplifier ensuring that the load and emitter resistor currents change together if the current I_{bias} is made constant, either because of the high supply voltage and correspondingly high resistance R, or by a separate low-voltage constant-current stage.

Fig. 1

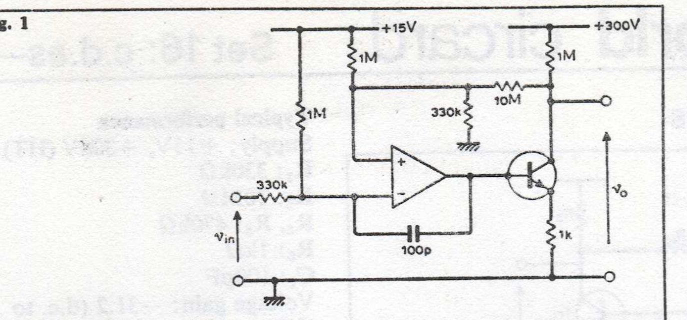


Fig. 2

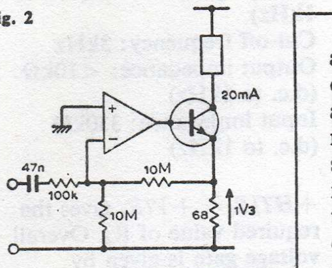
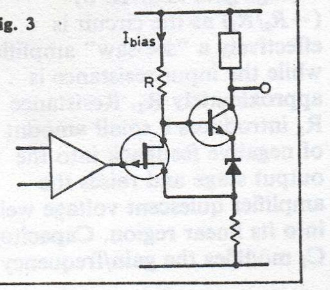


Fig. 3



Cross references
Set 7, cards 5, 7.
Set 16, cards 1, 2, 6.

controlled for inputs down to zero. This is best achieved for $R_1 = R_4$ when $V_o = V_i$ is the first-order approximation, the V_{be} effects at the two inputs cancelling.
R_4 As suggested, R_4 may equal R_1 for optimum stability but with gain restricted to +1. $R_4 \ngtr 2.0R_1$ except for a.c. amplifiers where some drift in quiescent conditions is acceptable.
R_5 1kΩ. Together with C_1 is part of the suggested network for maintaining stability. Limits amplifier current under load short circuit conditions providing protection for amplifier and output stage. Not adequate unless proper heat-sinking used since limit of output current depends on transistors' current-gains and is ill defined.
C_1, C_2 Control high frequency performance. C_1: 330p to 2.2nF, C_2: 5.6 to 22pF. Choose lowest values giving stability under operating conditions.
R_6 Load resistance. This may range down to 10Ω as currents of several amps are possible.

Circuit modifications

- A simpler circuit based on the same principles is Fig. M1. The limiting and compensation components have been eliminated together with the high-power transistor. The bias network shown is suitable for a source with resistive path to ground ≪470kΩ, with the non-inverting input grounding. The bias method is then basically the "nV_{be}" method as in the "amplified diode". Alternatively a capacitively coupled source may be used with no direct current in R_2 and $R_1 \approx 2R_3$ to set $V_o \approx +V/_2$.
- For higher efficiency the usual Class B technique can be applied with a complementary-symmetry output (Fig. M2). Not recommended for low-distortion applications, since additional bias networks to overcome crossover problems would lose the advantage of simplicity normally offered by this amplifier.
- Although the supply voltage capability is good (up to 36V) the output swing can be extended to ≈70V (Fig. M3) using a bridge configuration. The simplest arrangement for a.c. signals has the 1MΩ and 470kΩ resistors setting the output quiescent voltages to $+V/2$ for maximum undistorted voltage swing. Replacing the two input resistors by a potentiometer and applying the signal via the pot to opposite-phase inputs gives anti-phase outputs that can be set for equal magnitude with an overall voltage gain of about 18.

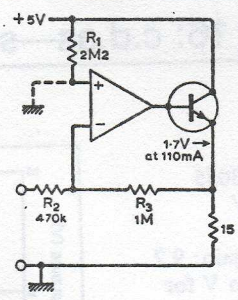


Fig M1

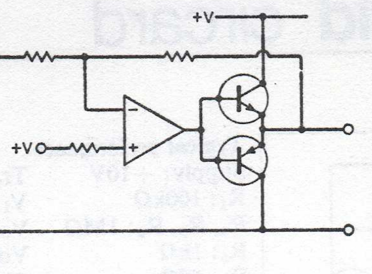
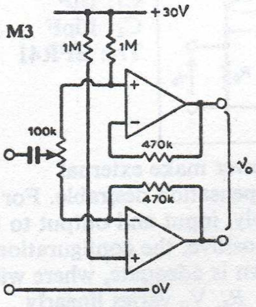


Fig. M2

Fig. M3



Cross references
Set 7, cards 2, 4, 7, 8, 10, 11.
Set 16, card 5.

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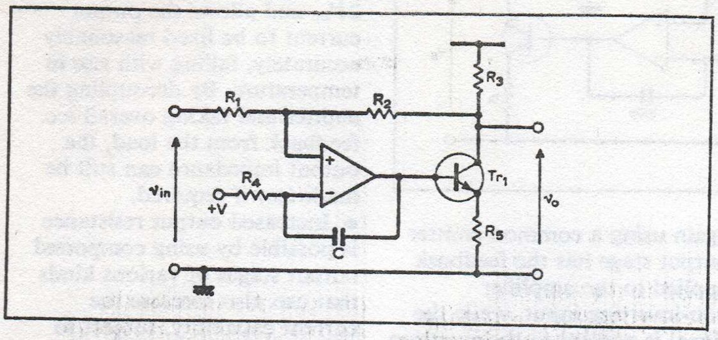
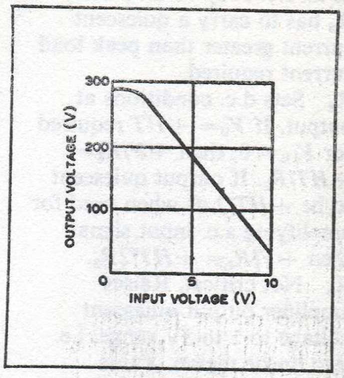
Set 16: c.d.as—signal processing—5

High voltage amplifiers



Typical performance
Supply: +15V, +300V (HT)
R_1: 330kΩ
R_2: 10MΩ
R_3, R_4: 470kΩ
R_5: 1kΩ
C_1: 100pF
Voltage gain: −31.2 (d.c. to 1kHz)
Cut-off frequency: 3kHz
Output impedance: <10kΩ (d.c. to 1kHz)
Input impedance: 330kΩ (d.c. to 1kHz)



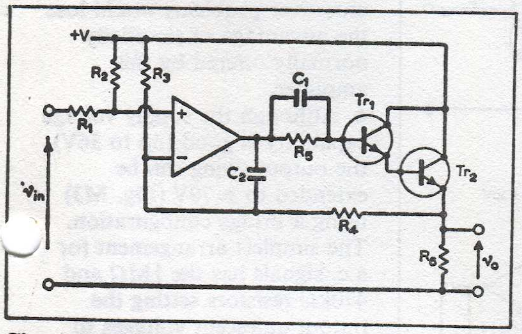
Circuit description
Where a high output voltage swing is required the amplifier must be fed from a separate low-voltage supply, and a suitable high-voltage transistor employed to withstand the main supply voltage. The configuration depends on whether it is the output voltage or current that is to be defined. If the former, then the feedback is taken from in shunt with the load. For an inverting-gain amplifier and the transistor in the common-emitter mode, the inverting gain of the transistor necessitates that the feedback be applied to what is normally considered as the non-inverting input. For an input of 0V d.c., the current flow in R_1 is small and that in R_2 is forced by the feedback to equal that in R_4. If in this condition the output is desired to be $+HT$ then $+HT/R_2 = +V/R_4$ gives the required value of R_4. Overall voltage gain is given by $(-R_2/R_1)$ as the circuit is effectively a "see-saw" amplifier while the input resistance is approximately R_1. Resistance R_5 introduces a small amount of negative feedback into the output stage and raises the amplifier quiescent voltage well into its linear region. Capacitor C_1 modifies the gain/frequency characteristic to maintain stability at the higher loop gain. Output voltage swing can be up to 95% of the supply voltage if lightly loaded and the negative feedback keeps the output impedance reasonably low.

Component changes
R_1, R_2 These set the input resistance and the voltage gain $R_2 \gg R_3$ to minimize loading on

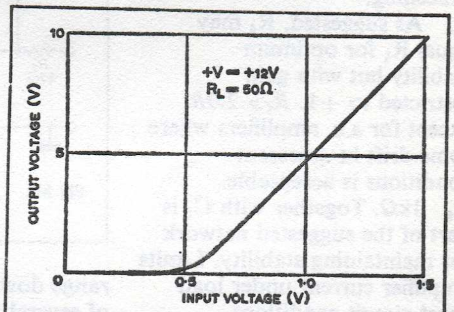
wireless world circard

Set 16: c.d.as—signal processing—6

Power amplifiers



Typical performance
Supply: +10V
R_1: 100kΩ
R_2, R_3, R_4: 1MΩ
R_5: 1kΩ
R_6: 50Ω
C_1: 1nF
C_2: 10pF
Tr_1: BFR41
Tr_2: TIP3055
V_i: 1.03V
V_o: 5.0V
Voltage gain: 9.7
Change in V for
I_L of 0-1A 0.05%
V_s of 10-14V 0.5%



Circuit description
Addition of a Darlington-connected pair of transistors increases the output current capability from 10mA to 1·5A depending on the ratings of the transistors used. One restriction is that the V_{be}'s of the transistors limit the output voltage to around 2.5V below the supply level allowing for the amplifier internal saturation. The additional phase shifts that may occur even in an emitter follower make external compensation desirable. For supply, input and output to be all positive, the configuration shown is adequate, where with $R_2 = R_3$, V_o varies linearly with V_i provided V_i is above the amplifier internal V_{be}. The relationship is then $V_o \approx V_{be} + (R_4/R_1)(V_i - V_{be})$ as the currents in R_1 and R_4 have to be equal. This means that as a d.c. amplifier it is of relatively low accuracy but is quite suitable for supplying small d.c. motors under the control of a phase-locked loop. A combination of the techniques for increasing the current ratings and voltage could allow the production of high-power amplifiers. Replacing the emitter follower stages by common emitter amplifier increases the available positive output swing to within a hundred millivolts of the positive supply.

Component changes
R_1 This sets the voltage gain in conjunction with R_4. Because of the V_{be} offset, the output voltage becomes temperature dependent particularly for V_i comparable to V_{be}, i.e. high gains are not compatible with good stability in this configuration. R_1 100k to 10MΩ.
R_2, R_3 These provide forward bias for each of the inputs allowing the output to be

the output. R_2 1M to 22MΩ.
R_3 Load resistance, dictated by user requirements. For use as an r.c.-coupled amplifier, R_3 has to carry a quiescent current greater than peak load current required.
R_4 Sets d.c. conditions at output. If $V_o = +HT$ required for $V_{in} = 0$, then $+V/R_4 = +HT/R_2$. If output quiescent to be $+HT/2$ as when used for amplifying a.c. input signal, then $+V/R_4 = +HT/2R_2$.
R_5 Not critical. Raises amplifier output quiescent voltage to 1 to 3V range, i.e. into linear region. Value dependent on output quiescent current but might be 50Ω to 5kΩ.
+V Chosen to suit amplifier, and available supplies (+4 to +36V).
+HT Dictated by load requirements.
Tr_1 Must have voltage rating in excess of +HT particularly if inductive loading possible.

Circuit modifications

- A non-inverting amplifier

Fig. 1

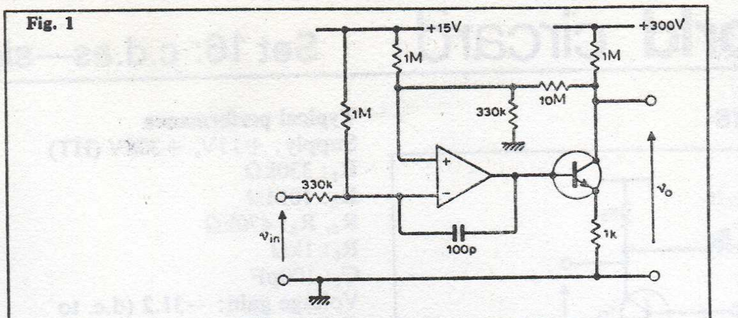
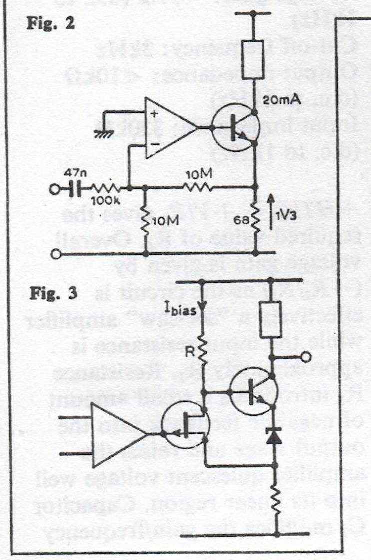


Fig. 2

Fig. 3



again using a common emitter output stage has the feedback applied to the amplifier non-inverting input, while the signal is applied to the inverting input (the inversion in the transistor ensuring the correct overall sign for the gain). For a transfer function that passes through the origin a pre-biasing network is required. Again the gain is set by the ratio of the feedback to the input resistor.

- For a high output-impedance, the feedback is taken from the transistor emitter. The method can be employed where the load is to be transformer coupled and the direct voltage drop across the primary is too small to allow of the d.c. feedback. The two 10-MΩ resistors define the potential at the transistor emitter as about $2V_{be}$ and allows the output current to be fixed reasonably accurately, falling with rise in temperature. By decoupling the emitter and taking overall a.c. feedback from the load, the output impedance can still be made low if required.
- Increased output resistance is possible by using compound output stages of various kinds that can also increase the current capability, subject to device power limitations. An f.e.t. draws no current from the amplifier ensuring that the load and emitter resistor currents change together if the current I_{bias} is made constant, either because of the high supply voltage and correspondingly high resistance R, or by a separate low-voltage constant-current stage.

Cross references
Set 7, cards 5, 7.
Set 16, cards 1, 2, 6.

controlled for inputs down to zero. This is best achieved for $R_1 = R_4$ when $V_o = V_i$ is the first-order approximation, the V_{be} effects at the two inputs cancelling.
R_4 As suggested, R_4 may equal R_1 for optimum stability but with gain restricted to +1. $R_4 \ngtr 2.0R_1$ except for a.c. amplifiers where some drift in quiescent conditions is acceptable.
R_5 1kΩ. Together with C_1 is part of the suggested network for maintaining stability. Limits amplifier current under load short circuit conditions providing protection for amplifier and output stage. Not adequate unless proper heat-sinking used since limit of output current depends on transistors' current-gains and is ill defined.
C_1, C_2 Control high frequency performance. C_1: 330p to 2.2nF, C_2: 5.6 to 22pF. Choose lowest values giving stability under operating conditions.
R_6 Load resistance. This may range down to 10Ω as currents of several amps are possible.

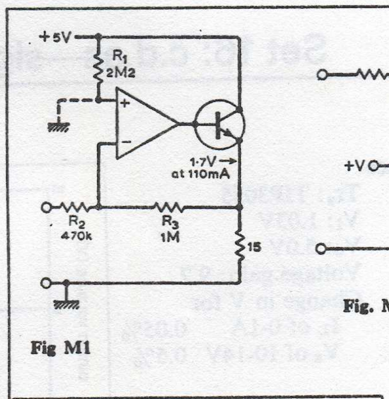


Fig M1

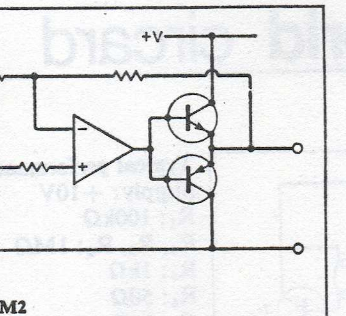
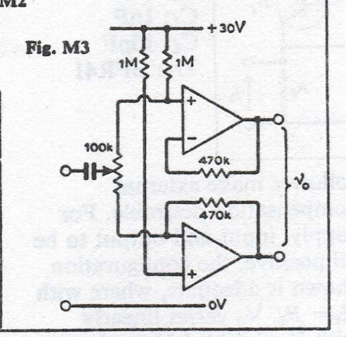


Fig. M2

Fig. M3

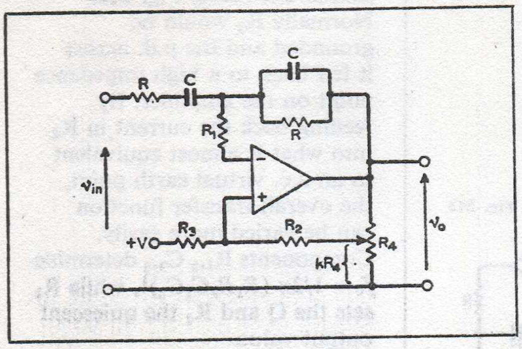


Circuit modifications

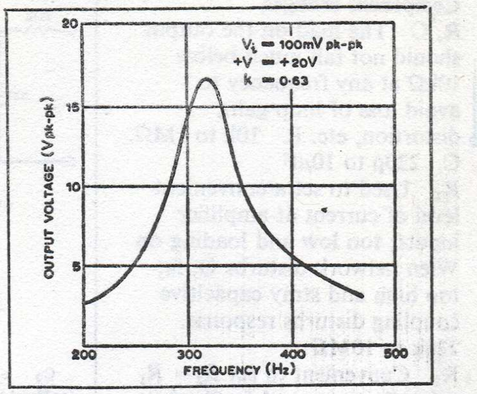
- A simpler circuit based on the same principles is Fig. M1. The limiting and compensation components have been eliminated together with the high-power transistor. The bias network shown is suitable for a source with resistive path to ground ≤470kΩ, with the non-inverting input grounding. The bias method is then basically the "nV_{be}" method as in the "amplified diode". Alternatively a capacitively coupled source may be used with no direct current in R_2 and $R_1 \approx 2R_3$ to set $V_o \approx +V/_2$.
- For higher efficiency the usual Class B technique can be applied with a complementary-symmetry output (Fig. M2). Not recommended for low-distortion applications, since additional bias networks to overcome crossover problems would lose the advantage of simplicity normally offered by this amplifier.
- Although the supply voltage capability is good (up to 36V) the output swing can be extended to ≈70V (Fig. M3) using a bridge configuration. The simplest arrangement for a.c. signals has the 1MΩ and 470kΩ resistors setting the output quiescent voltages to $+V/2$ for maximum undistorted voltage swing. Replacing the two input resistors by a potentiometer and applying the signal via the pot to opposite-phase inputs gives anti-phase outputs that can be set for equal magnitude with an overall voltage gain of about 18.

Cross references
Set 7, cards 2, 4, 7, 8, 10, 11.
Set 16, card 5.

Bandpass filters

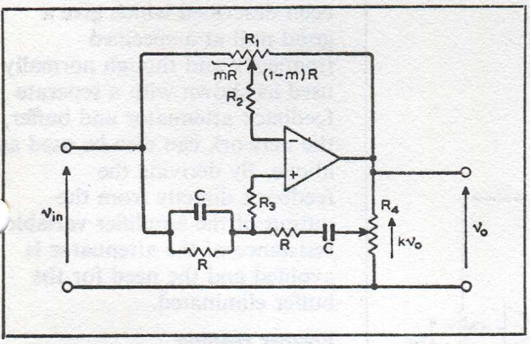


Typical performance
R: 47kΩ
C: 10nF
R_1, R_2: 1MΩ
R_3: 6.8MΩ
R_4: 100kΩ
Supply: +15V
f_0: 320Hz
For $Q = 15$
$k = 0.640$
N.B. Onset of oscillation at $k = 0.648$ compared with theoretical value of 0.66.

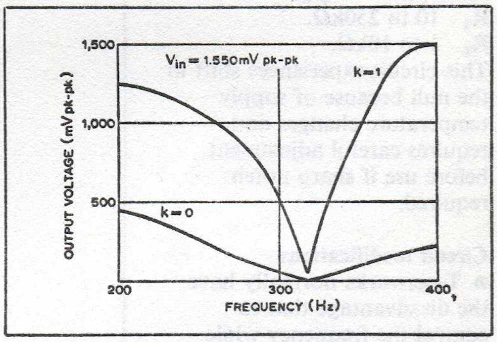


Circuit description
Active filter techniques based on operational amplifiers are applicable to current-differencing circuits. That shown is a direct adaptation of Circard 1, Set 1. A Wien network is placed between source and output with the potential at its junction monitored by the amplifier input terminal. Resistor R_1 is a high-value resistance to minimize loading on the network. Resistor R_2 provides a variable amount of positive feedback to increase the Q of the circuit with minimal effect on the centre-frequency while R_3 sets the quiescent output voltage to allow for maximum signal swing. Because $R_2 > R_1$ would be required for stability, with R_2 taken directly to the output, and variable high-value resistors are inconvenient, the alternative is to tap R_2 onto a variable portion of the output voltage. The result is a band-pass filter with centre frequency given by $f = 1/2\pi CR$ and with Q controlled by R_4. The limited gain and bandwidth capabilities of the amplifier does not allow the circuit to provide large stable Qs nor to operate successfully at frequencies ⩾10kHz. The input impedance falls as the Q is increased because of the increased output swing (gain ∝ Q to a first order). It is also true that the sensitivity of such circuits to variations in component values is proportional to the Q, i.e. for $Q = 10$ a 1% change in a critical resistor might

Notch filters



Typical performance
Supply: +20V
R_1: 100kΩ
R_2, R_3: 1M
R_4: 1kΩ
R: 47kΩ
C: 10nF
Notch frequency: 319Hz
Achieved for $m \approx 0.35$
for $k = 0$ and $k = 1$



Circuit description
The circuit is again derived from a Wien Bridge to demonstrate the principles by which known circuits can be adapted to current-differencing amplifiers. It provides a notch or null in the response at a frequency set by the RC values ($f = 1/2\pi RC$) with R_1 providing a trimming action to get as true a null as possible. If R_4 is a low resistance potentiometer, then the depth of the null is unaffected as the tapping point is varied, since both ends of the potentiometer are at zero for this condition. However, the positive feedback introduced has the effect off-null of sharpening up the response so that the gain remains close to unity for frequencies close to the null. This makes the circuit useful for nulling out the fundamental (or particular harmonic) of a complex waveform with minimal effect on the other harmonics. A weakness of this particular circuit is that it relies on the matching of the inverting and non inverting input sensitivities making it prone to variation with supply/temperature changes, etc.

Component changes
R, C The difficulty with this particular circuit is that these should have a low impedance compared with R_3 so that the latter does not load them, while not in turn being disturbed by the varying source resistance of R_4 as k is varied. Typically R 10 to 100kΩ

produce >10% variation in the Q itself.

Component changes
R, C The load on the output should not fall much below 10kΩ at any frequency to avoid loss of loop gain, distortion, etc. R 10k to 1MΩ, C 220p to 10μF
R_1 Used to set a convenient level of current at amplifier inputs; too low and loading on Wien network disturbs Q, f_0; too high and stray capacitive coupling disturbs response. 220k to 10MΩ.
R_2 Convenient to set $R_2 = R_1$ using R_4 to control feedback. Alternatively for fixed Q particularly if low, eliminate R_4 taking R_2 directly to output. Then $R_2 > 1.5R_1$.
R_4 Not critical, but $\gg$10kΩ and $< R_2$.
R_3 Sets quiescent output voltage for maximum swing. Should be chosen after other components have been selected for desired response. Typically 2 to 10 × R_1, but for a given value, small variations

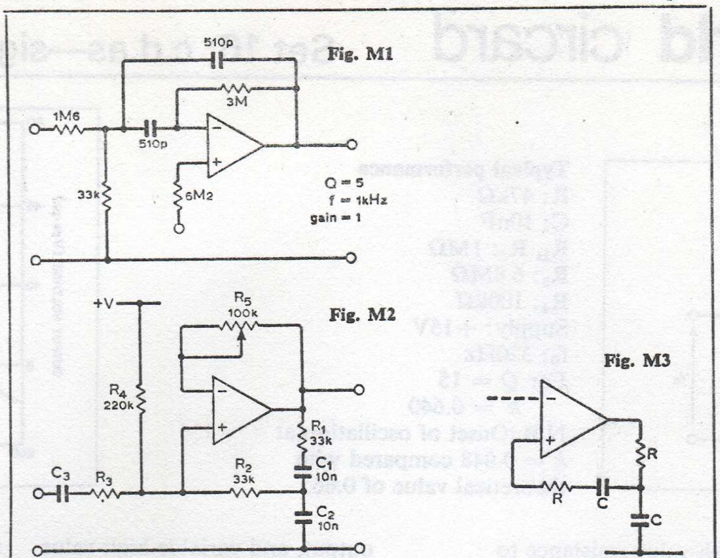


Fig. M1

Fig. M2

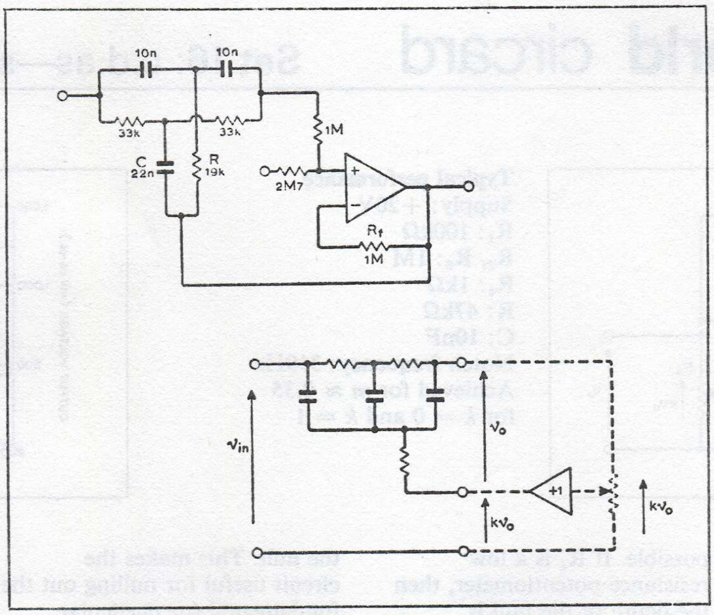
Fig. M3

in k and hence Q can be accommodated without serious bias changes.

Circuit modifications
● Multiple feedback circuits (Fig. M1) are the equivalent of the virtual earth circuits used with conventional operational amplifiers. With correct scaling of resistors, capacitors $Q > 1$ is achieved simultaneously with centre-frequency gain of unity. Again both are strongly dependent on component stability at high-Q values.
● A better approach to filter design with current-differencing amplifiers is to select passive networks whose transfer-function gives a defined output current. It is possible to adapt the Wien and similar networks, Fig. M2. Normally R_3 would be grounded and the p.d. across it fed back to a high impedance point on the amplifier. By feeding back the current in R_2 into what is almost equivalent to an a.c. virtual earth point, the overall transfer function can be varied more easily. Components $R_{1,2}$ $C_{1,2}$ determine $f_0 = 1/2\pi\,(R_1R_2C_1C_2)^{\frac{1}{2}}$, while R_5 sets the Q and R_4 the quiescent output value.
● Any of the other RC networks used in bandpass filter designs are applicable and Fig. M3 gives almost identical performance for $R_1 = R_2 = R$, $C_1 = C_2 = C$.

Cross references
Set 1, cards 1, 3, 6, 7, 8, 12.
Set 5, cards 6, 10.
Set 10, card 8.
Set 16, cards 8, 9.

with C to give frequency as $f = 1/2\pi RC$ C, 220p to 10μF.
R_2, R_3 1 to 10MΩ. At higher frequencies stray capacitance effects bring phase shifts preventing good null.
R_1 10 to 250kΩ.
R_4 1 to 10kΩ.
This circuit experiences shift in the null because of supply temperature changes and requires careful adjustment before use if sharp notch required.



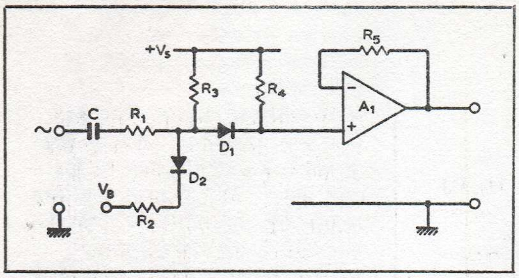
Circuit modifications
● T-networks normally have the disadvantage that to control the frequency while retaining a sharp notch requires adjustment of a number of interacting components. Interconnected with an amplifier (top circuit), using positive feedback to steepen the sides of the notch (equivalent to varying the Q but not the centre frequency), it has the advantage that variations in the amplifier gain only affect the steepness leaving the depth unaffected. A further advantage accrues using the current-differencing amplifier in that the gain can be adjusted, while retaining a fixed input impedance, by varying R_f. This also affects the gain at frequencies well away from the notch, but the approach is often used where the gain is just less than unity the sides of the notch can be very steep and unity gain can be assumed off resonance. Independent adjustment of C and R are required to obtain a complete null and the system is easier to use if fine frequency control is possible at the oscillator, as may be possible when used as part of a distortion measuring system.
● Other passive networks have been described which give a good null at a specified frequency and though normally used as shown with a separate feedback attenuator and buffer, the network can also be used as above. By deriving the feedback directly from the output of the amplifier variable resistance of the attenuator is avoided and the need for the buffer eliminated.

Further reading
Rowe, N. B. Designing a low frequency active notch filter, *Electronic Engineering*, April 1972.
Dance, J. L. & Edwards, K. H. Simple null filter with variable notch frequency, *Electronic Engineering*, vol. 36 1964, pp.478/9.

Cross references
Set 1, cards 9, 10.
Set 16, cards 7, 9.

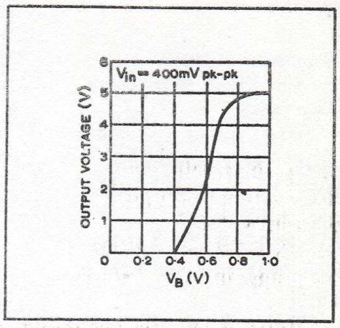
wireless world circard Set 16: c.d.as—signal processing—10

Gain-controlled amplifiers



Typical performance

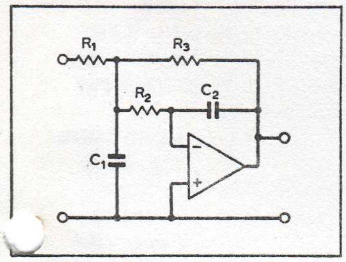
Supply	+15V
A_1	¼LM3900
R_1	2.2kΩ
R_2	100Ω
R_3,R_4	100kΩ
R_5	33kΩ
C_1	10μF
V_{in}	400mV pk–pk 10kHz
for V_B	0.9V
V_o	5V pk–pk



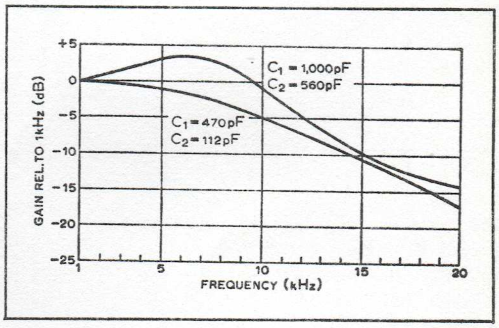
Circuit description
The non-inverting input has a diode-connected transistor as part of a current-mirror. For a given feedback resistor to the inverting input, the current flowing in the non-inverting input gives a proportional output voltage. The non-linearity of the input impedance leads to output waveform distortion if any low-resistance parallel path is used to attenuate the signal. If the path consists of a suitably-biased silicon diode then its non-linearity is comparable with that of the input stage and the input signal division between the two paths has little variation with signal amplitude, i.e. little distortion results. This remains true while the signal current is well below the bias level.
As the direct current in the diode is increased, the corresponding fall in slope-resistance by-passes more signal current reducing the gain. Resistor R_4 provides sufficient bias current to set the output potential to a minimum of $V_s/3$. This applies when the bias voltage is low with the direct current in R_3 as well as the a.c. signal current via R_1 are both shunted away through R_2, D_2. The gain is virtually zero in this condition. As V_B approaches one diode voltage, the direct currents in D_1, D_2 and the amplifier input become comparable. The slope resistances are also comparable and the input current divides equally between the amplifier and D_2. For any further increase in V_B the signal current flows on through D_1 to the amplifier with negligible attenuation by D_2. At this extreme the direct current in R_3 contributes to the amplifier bias and the output d.c. level rises to $2V_s/3$.

wireless world circard Set 16: c.d.as—signal processing—9

Low pass/high pass filters



Typical performance
Supply: 5V
R_1, R_3: 150kΩ
R_2: 47kΩ
C_1: 470pF
C_2: 120pF
−3dB point ≈ 8kHz (low frequency gain about −1)
Max. input/output swing ≈ 1V pk-pk.



Circuit description
The most convenient configuration for most feedback circuits with these amplifiers is the virtual earth configuration; it is not possible to use series applied feedback as with standard Sallen & Key type filters.
The network is not easy to analyse component-by-component because of the interactions between them but the overall transfer function is well defined particularly if the gain in the pass-band is low, e.g. unity. Then it is sufficient if the amplifier voltage gain is >100 to have a transfer function that is very close to the theoretical value. A second-order low-pass filter results where the cut-off frequency may range from 1Hz to >100kHz. A convenient means of adjustment of the filter properties is via C_1, C_2. It is their product that fixed the cut-off frequency (for R_1, R_2, R_3 fixed) while their ratio determines the shape of the transfer function (i.e. Butterworth, etc.). The output impedance is low and the input impedance may be up to 1MΩ if required allowing such filters to be cascaded with negligible loading.
As shown the output voltage is defined as $(R_3/R_1 + 1)V_{be}$ provided the input has a quiescent value of zero, as would be achieved by capacitive coupling from the previous stage. Since $R_1 = R_3$ is a convenient value for design of the filter characteristics this restricts the output to a quiescent value of about 1.2V regardless of the supply. This may be inconvenient where larger voltage swings are desired and alternative biasing schemes may be needed (see Circuit modifications).